TRANSISTOR CIRCUIT CHARACTERISTICS

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by

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PREFACE

This paper was done in the spring of 1953 at the United States Naval Postgraduate School. The purpose was to search the tremendous volume of transistor literature and to extract information on the circuit characteristics in those particular applications which seemed most promising. An effort was made to select the most generally accepted methods of utilizing the known physical characteristics. After the material was selected it was expanded or compressed, as appropriate, to give a practical engineering point of view.

The writer wishes to acknowledge, with appreciation, the assistance of Professors A. Sheingold, and D. A. Stentz in selecting and arranging the material, and in correcting the manuscript.

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LIST OF ILLUSTRATIONS

TABLE OF SYMBOLS AND ABBREVIATIONS

Symbols	Name
PCT	Point Contact Transistor(s)
JT	Junction Transistor(s)
α	current gain of transistor
R_g	Generator internal resistance
R_L	Load Resistance
R_{ij}	General network resistance parameter
R^*_{ij}	Transistor four terminal open circuit resistance parameter
Δ	Value of a determinant
r	small signal equivalent circuit parameter
I	Direct current
i	Alternating Current
g_m	Vacuum tube transconductance
e	Alternating voltage
R_o	Output resistance
R_{in}	Input resistance
K	Voltage Gain
G	Power Gain
P	Power
V	Voltage Drop DC
E	Voltage source DC
S	Point spacing of PCT

ρ	Resistivity
μ	Mobility of minority carrier
σ	Conductivity
f_c	Cut-off frequency
T	Transit time

CHAPTER I

ABSTRACT

The purpose of this paper is first, to show how transistor small signal equivalent circuits can be deduced from either their static characteristics or open circuit resistance measurements, and then, by the use of circuit theory, to utilize these equivalent circuits to predict the characteristics of some typical amplifier circuits.

By-products of this development are the stability criterion for amplifiers and the negative resistance characteristics of transistors.

The latter by-product is the subject of a fairly detailed presentation of a useful method of analyzing negative resistance circuits by subdividing the negative resistance characteristic into three approximately linear regions.

In the interests of completeness there are included some general comments on frequency response, large signal analysis, noise, temperature effects, and, finally, a representative collection of practical circuits with brief comments on the outstanding features of each.

It is hoped that this material will be useful in bridging the gap between the theoretical and the practical aspects of transistor circuits.

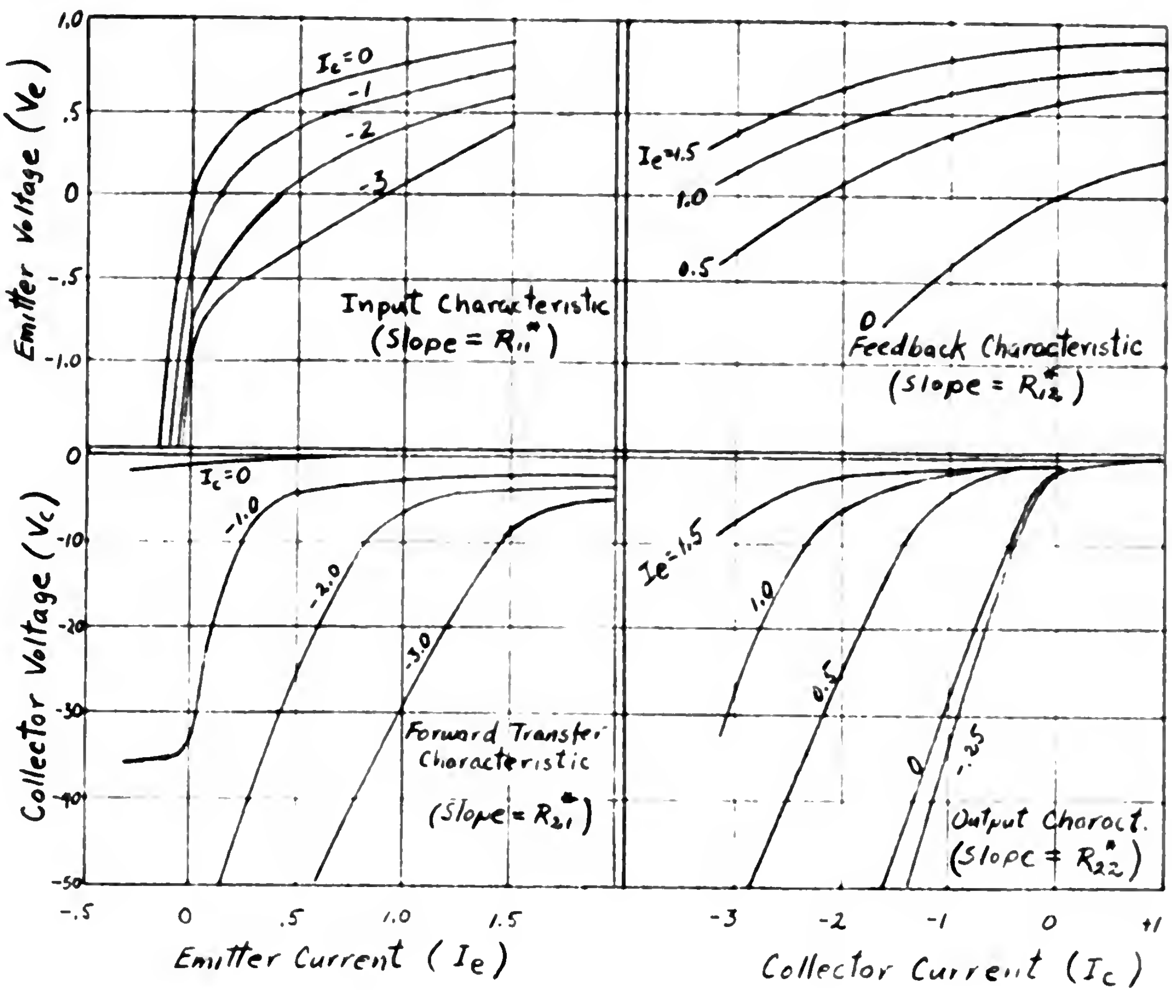
The material contained herein was derived primarily from periodical literature.

CHAPTER II

AC EQUIVALENT CIRCUIT

1. Static Characteristics,[2,9,14,22].

The static characteristics of transistors are used in essentially the same way as those of vacuum tubes[14]; the small signal parameters of the small signal equivalent circuits are obtained from the slopes of the curves in a manner that is analogous[2]. A typical set of static characteristics of a PCT is shown below[9].



The slopes of the curves above, at the chosen operating point, give the open circuit resistances indicated on each curve[1,2,9,10,14,22]. These important parameters may also be obtained by direct open circuit measurement[2,17]. In making these measurements current is used as the independent variable since, under conditions to be mentioned later, current at any pair of terminals may be a multiple valued function of an independent voltage variable[2,14]. The open circuit resistances can be used to obtain several equivalent circuits. The most convenient of these is a simple T[2,10,14].

2. Derivation of Equivalent T.

Before proceeding with the derivation of the equivalent T it should be mentioned that a fairly recent paper[17] has shown that the transistor may be handled as a four terminal "black box" using newly developed active circuit matrix methods. If matrix methods are used it is not necessary to interpret the results of the open circuit measurements but if an equivalent circuit is to be deduced from the measurements it is necessary to conceive a physically realizable circuit to represent the results of the measurements. We shall therefore interpret the open circuit resistances (R*) in the light of the fact that there is known to be current gain in the forward direction[9,14,21,22,27].

R^*_{12} is not equal to R^*_{21} (2,14,22). The mutual resistance, R^*_{12} is a passive resistance common to the input and output circuits. R^*_{21} must include this common resistance, but the amount by which R^*_{21} exceeds R^*_{12} is a result of the additional effect of current in the input circuit on the output; i.e., there is an effect over and above that chargeable to direct mutual resistance coupling. We will designate this forward transfer resistance r_m and the passive mutual resistance r_b. Then, as stated above,

$$R^*_{12} = r_b$$
$$R^*_{21} = r_b + r_m \tag{1}$$

Now the difference between R^*_{11} and R^*_{12} must be the remaining passive resistance of the input circuit when the output is open circuited. This is the emitter resistance, r_e. Similarly the difference between R^*_{22} and r_b is the collector resistance, r_c.

Summarizing:

$$R^*_{11} = r_e + r_b \qquad R^*_{12} = r_b$$
$$R^*_{21} = r_m + r_b \qquad R^*_{22} = r_c + r_b \tag{2}$$

If neither R^*_{11} nor R^*_{22} is to include r_m we must represent the active unilateral property by an equivalent voltage or current generator.

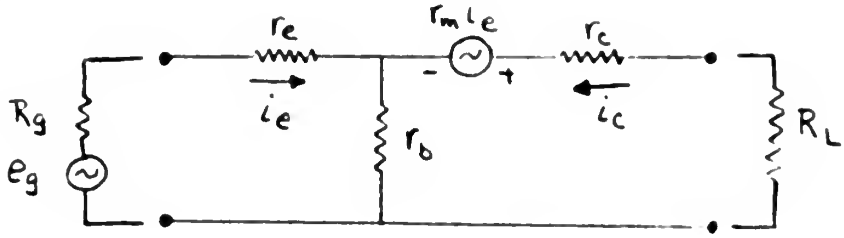
r_m obviously expresses a ratio of a voltage to a current. Specifically it is a measure of the voltage in the collector circuit which results from an increment of I_e (i_e) exclusive of that caused by the direct coupling, r_b.

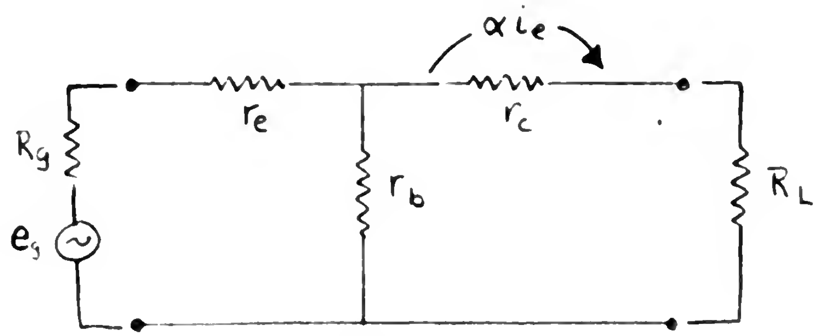
$V = r_m i_e$ is the voltage generator(14,27,28). This is similar to $g_m e_g$, the current generator in the small signal equivalent circuit of a vacuum tube. As is the case with vacuum tubes, this might be replaced by its dual, in this case, a current generator(28).

The ratio of the increment of current in the collector circuit to the increment of emitter current causing it has been called the current gain and designated α (21,22,27,29). Therefore, the equivalent current generator is αi_e.

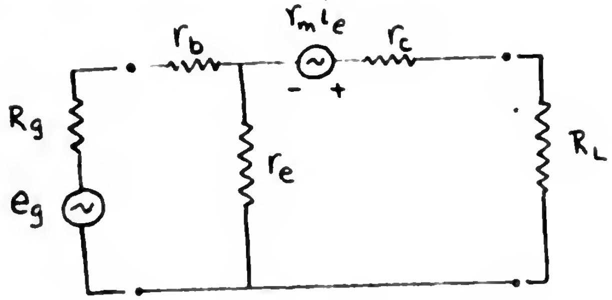
3. Amplifier Equivalent Circuits.

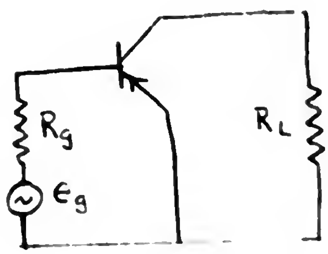
For the grounded base connection, the two ac equivalent circuits are(10,14,22,27):

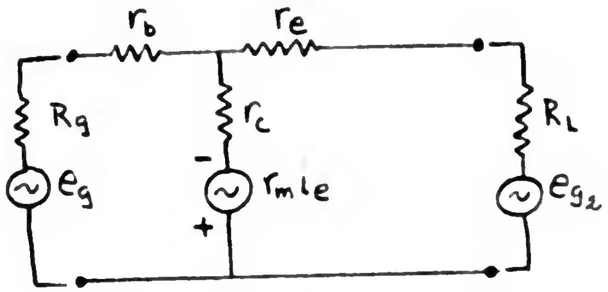


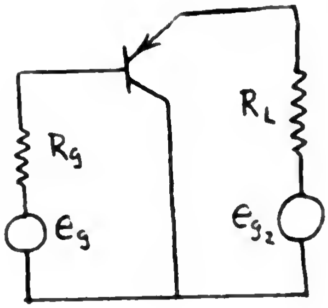


Similarly, the voltage generator equivalent circuits for grounded emitter and collector are respectively[14,22]:









The reason for the inclusion of the generator, in what is normally the output of the last circuit above, is that under the condition that $\alpha > 1$ there is appreciable gain in the reverse direction[14,22]. In fact, when $\alpha = 2$ the forwarded and reverse power gains are equal[14,22].

CHAPTER III

INPUT AND OUTPUT IMPEDANCES OF AMPLIFIERS[2,12,14]

1. General.

We will now develop the input and output impedances of simple amplifier circuits from the foregoing equivalent circuits.

For clarity a numerical example will be carried along with the literal. Given:

$R_{11}^* = 530\Omega$	$R_g = 500\Omega$	$r_e = 240\Omega$
$R_{12}^* = 290\Omega$	$R_L = 20{,}000\Omega$	$r_b = 290\Omega$
$R_{21}^* = 34{,}000\Omega$		$r_c = 18{,}710\Omega$
$R_{22}^* = 19{,}000\Omega$		$r_m = 33{,}710\Omega$

The general circuit determinant for all three cases is[5,14]:

$$\Delta = \begin{vmatrix} R_{11} & R_{12} \\ R_{21} & R_{22} \end{vmatrix} = R_{11}R_{22} - R_{12}R_{21} \qquad (3)$$

From four terminal network theory the input impedance, as seen from the generator terminals, is[5]

(4)

$$R'_{in} = \frac{\Delta}{A_{11}}$$

, where A is the cofactor of the first element, a_{11}, of the determinant

therefore,

$$R'_{in} = R_{11} - \frac{R_{12}R_{21}}{R_{22}} \qquad (5)$$

or, letting R_g be the internal impedance of the generator, the impedance looking into the input terminals becomes,

$$R_{in} = R_{11} - R_g - \frac{R_{12}R_{21}}{R_{22}} \qquad (6)$$

In a similar way the output impedance seen by the load is,

$$R_o = R_{22} - R_L - \frac{R_{12} R_{21}}{R_{11}} \qquad (7)$$

2. Grounded Base.

Applying the general formulae to the grounded base connection we have,

$$R_{in} = r_e + r_b - \frac{r_b(r_b + r_m)}{r_c + r_b + R_L} = 530 - \frac{290 \times 34000}{39000} = 277\Omega \qquad (8)$$

$$R_o = r_c + r_b - \frac{r_b(r_b + r_m)}{r_e + r_b + R_g} = 19000 - \frac{290 \times 34000}{1030} = 9420\Omega \qquad (9)$$

3. Grounded Emitter.

$$R_{in} = r_e + r_b + \frac{r_e(r_m - r_e)}{r_e + r_c + R_L - r_m} = 530 + \frac{290 \times 33470}{5240} = 2385\Omega \qquad (10)$$

$$R_o = r_e + r_c - r_m + \frac{r_e(r_m - r_e)}{r_e + r_b + R_g} = -14760 + \frac{290 \times 33470}{1030} = -5530\Omega \qquad (11)$$

4. Grounded Collector.

$$R_{in} = r_b + r_c - \frac{r_c(r_c - r_m)}{R_L + r_e + r_c - r_m} = 19000 + \frac{18710 \times 15000}{5340} = 72600\Omega \qquad (12)$$

$$R_o = r_e + r_c - r_m + \frac{r_c(r_m - r_c)}{R_g + r_b + r_c} = -14760 + \frac{18710 \times 15000}{19500} = -1660\Omega \qquad (13)$$

CHAPTER IV

VOLTAGE AND POWER GAINS (2,12,14,17,27)

Gain formulae will now be developed for the same three connections carrying through with the same numerical constants.

1. Grounded Base.

$$R_{11} = R_{11}^{*} + R_g \qquad R_{12} = R_{12}^{*}$$

$$R_{21} = R_{21}^{*} \qquad R_{22} = R_{22}^{*} + R_L \tag{14}$$

$$\Delta = \begin{vmatrix} R_{11} & R_{12} \\ R_{21} & R_{22} \end{vmatrix} = \begin{vmatrix} 1030 & 290 \\ 34000 & 39000 \end{vmatrix} = 30.31 \times 10^6$$

Referring to the equivalent circuit it may be shown that

$$i_L = - \frac{R_{21}\, e_g}{\Delta} \tag{15}$$

then,

$$e_L = -i_L R_L = \frac{R_{21} R_L\, e_g}{\Delta} \tag{16}$$

hence,

$$\frac{e_L}{e_g} = \frac{R_{21} R_L}{\Delta} \tag{17}$$

Therefore the forward voltage gain is

$$K_F = \frac{R_{21} R_L}{\Delta} = \frac{34 \times 20}{30.31} = 22.4 \tag{18}$$

The power available from the source is

$$P_{in} = \frac{e_g^{2}}{4 R_g} \tag{19}$$

and the power consumed in the load is

$$P_o = \frac{e_L^2}{R_L} \tag{20}$$

The forward power gain is therefore,

$$G_F = \frac{P_o}{P_{in}} = \frac{e_L^2/R_L}{e_g^2/4R_g} = \frac{4R_g}{R_L}\left(\frac{e_L}{e_g}\right)^2 = \frac{4R_g}{R_L} K_F^2 \tag{21}$$

$$= \frac{4R_g}{R_L}\left(\frac{R_{21}R_L}{\Delta}\right)^2 = 4R_g R_L\left(\frac{R_{21}}{\Delta}\right)^2 = 50.4 \tag{22}$$

The reverse gains are obtained in a similar manner, as follows:

$$i_g = -\frac{e_{g_2} R_{12}}{\Delta} \tag{23}$$

$$e_{L_1} = -i_g R_g = \frac{R_g R_{12} e_{g_2}}{\Delta} \tag{24}$$

$$K_R = \frac{e_{L_1}}{e_{g_2}} = \frac{R_g R_{12}}{\Delta} = \frac{500 \times 290}{30.31 \times 10^6} = .00479 \tag{25}$$

$$G_R = \frac{4R_L}{R_g} K_R^2 = 4R_g R_L\left(\frac{R_{12}}{\Delta}\right)^2 \approx 0 \tag{26}$$

2. Grounded Emitter.

$R_{11} = r_e + R_g = 1030 \qquad R_{12} = r_e = 290$

$R_{21} = r_e - r_m = -33470 \qquad R_{22} = r_e + r_c + R_L - r_m = 5240$

$$\Delta = \begin{vmatrix} 1030 & 290 \\ -33470 & 5240 \end{vmatrix} = 15.1 \times 10^6 \qquad (27)$$

$$K_F = \frac{R_{21} R_L}{\Delta} = \frac{-33470 \times 20000}{15.1 \times 10^6} = -44.7 \qquad (28)$$

$$G_F = \frac{4 R_g}{R_L} K_F^2 = 200 = +23\,db \qquad (29)$$

$$K_R = \frac{R_g R_{12}}{\Delta} = \frac{500 \times 290}{15.1 \times 10^6} = .0096 \qquad (30)$$

$$G_R = \frac{4 R_L}{R_g} K_R^2 = 4 R_g R_L \left(\frac{R_{12}}{\Delta}\right)^2 = 4 R_g R_L \left(\frac{r_e}{\Delta}\right)^2 \qquad (31)$$

$$= 10.24 \times 10^{-3} = .01024 = -19.9\,db$$

3. Grounded Collector.

$$R_{11} = r_b + r_c + R_g = 19500 \qquad R_{12} = r_c - r_m = -15000$$

$$R_{21} = r_c = 18710 \qquad R_{22} = r_e + r_c + R_L - r_m = 5240$$

$$\Delta = \begin{vmatrix} 19500 & -15000 \\ 18710 & 5240 \end{vmatrix} = 383 \times 10^6 \tag{32}$$

$$K_F = \frac{R_{21} R_L}{\Delta} = \frac{18710 \times 20000}{383 \times 10^6} = .978 \tag{33}$$

$$G_F = \frac{4 R_g}{R_L} K_F^2 = .0958 = -10.18 \text{ db} \tag{34}$$

$$K_R = \frac{R_g R_{12}}{\Delta} = \frac{500(-15000)}{383 \times 10^6} = -.0197 \tag{35}$$

$$G_R = \frac{4 R_L}{R_g} K_R^2 = 160 \times 3.89 \times 10^{-4} \tag{36}$$

4. Summary.

In the three cases just developed the values of R_g and R_L were the same (500 Ω and 20 K respectively).

Summarizing the input and output impedances of the simple amplifiers:

Grounded base

$$R_{in} = 277\,\Omega$$

$$R_o = 9420\,\Omega$$

Grounded emitter

$$R_{in} = 2.385 \text{ K}$$

$$R_o = -5.33 \text{ K}$$

Grounded collector

$$R_{in} = 72.6 \text{ K}$$

$$R_o = -1.66 \text{ K}$$

It is apparent that the chosen values of R_g and R_L are very inappropriate for the last two connections. In order to reveal an interesting feature of the grounded collector amplifier let us choose R_g equal 20 K and R_L equal 20K, then

$$R_{11} = 39K \qquad R_{12} = -15K$$

$$R_{21} = 18.71K \qquad R_{22} = 5.24K$$

$$\Delta = (39 \times 5.24 + 15 \times 18.71)10^6 = 76.5 \times 10^6 \tag{37}$$

and

$$K_F = \frac{18.71 \times 20}{76.5} = 4.9 \tag{38}$$

and $$K_R = \frac{20(-15)}{76.5} = -3.92 \tag{39}$$

Here it is seen that K_F and K_R are of the same order of magnitude but of opposite sign. In other words the grounded collector connection offers the possibility of use as a bilateral amplifier with a phase reversal in the reverse but not in the forward direction. The grounded emitter offers similar possibilities but to a lesser extent.

CHAPTER V

STABILITY CRITERION[9,13,14,21,22]

Any circuit which has a negative circuit determinant is unstable[22]. This is a mathematical representation of the condition in which the feedback is greater than that necessary to make the loop gain unity. In terms of the usual knowns of a transistor circuit this criterion may be roughly reduced to give the conditions conducive to instability.

(1) Large feedback resistance

(2) Large α

(3) Small R_g

(4) Small R_L

If $\alpha < 1$ for either PCT of JT and there are no external feedback resistances, then the transistor is unconditionally stable[9,13,21,22]. PCT in general have current gain greater than unity[9,21]; hence, they may be short circuit unstable if the internal feedback resistance, r_b is sufficiently large.

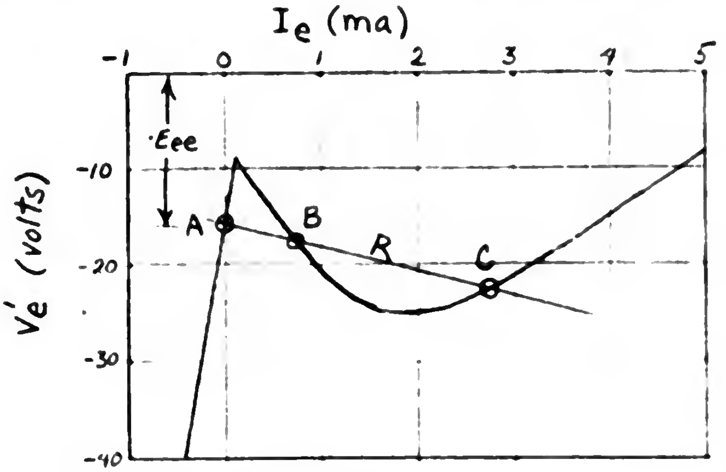
It should be mentioned here that a positive value of Δ is not an assurance of stability but that a negative value guarantees instability.

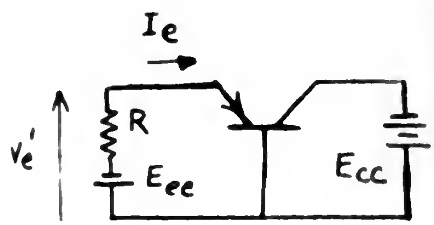
CHAPTER VI

NEGATIVE RESISTANCE CHARACTERISTICS[1,4,22]

1. General.

As mentioned previously, under certain conditions, i.e., α and r_b large, any two terminals of a transistor may exhibit a negative resistance characteristic[9]. In these cases the voltage vs current curves will show current to be a multiple valued function of voltage or vice versa. This is the case, for example, when a transistor which is short circuit unstable has its collector short circuited through a collector bias battery. Then the emitter characteristic might appear as shown below[1,4,9].



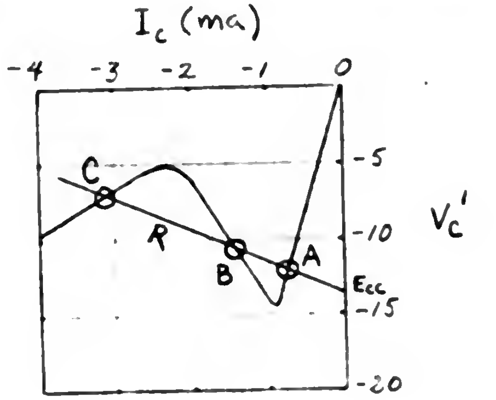


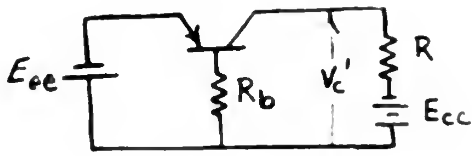
In this example the circuit has two stable points, A and C. If operating at point A and a positive pulse is applied to the emitter, operation will shift practically instantaneously[1,4] to point C provided that the amplitude of the pulse is sufficient to attain the negative resistance

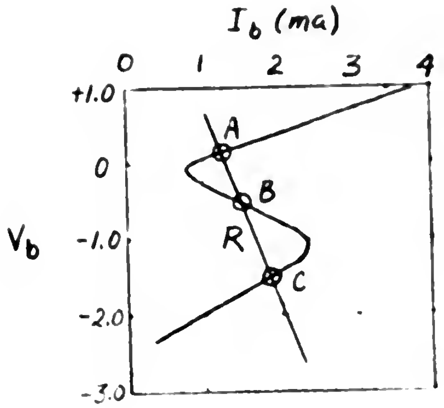
portion of the curve. To return to A a negative pulse is required.

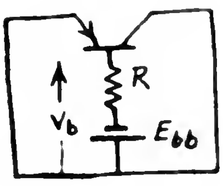
In practical circuits the transistor need not be short circuit unstable, for it is made so by the inclusion of a suitably large external base resistance, R_b [1].

The negative resistance curves for collector and base with load lines shown for bi-stable operation are given below[1,4]. These curves are somewhat idealized.





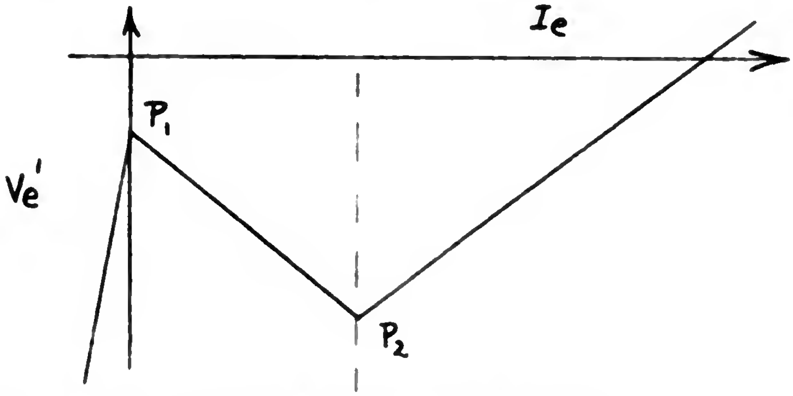




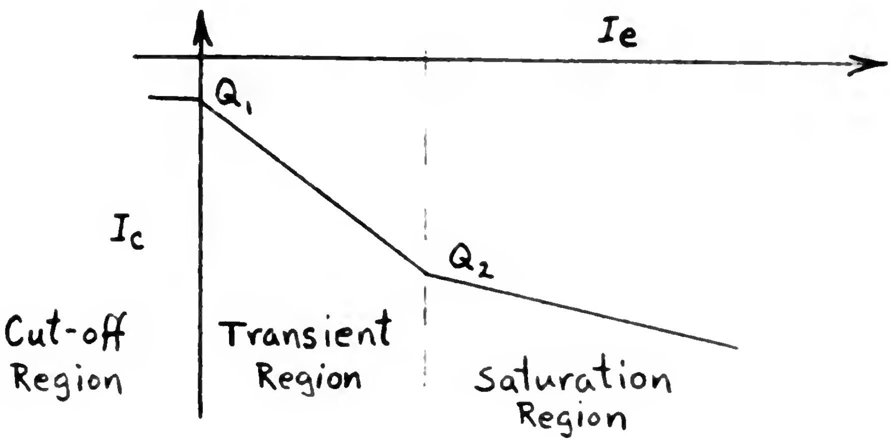
2. Linear Segment Analysis (general)[1,4,7].

For purposes of analysis let us take the negative resistance emitter circuit and the associated emitter input and current transfer characteristics, and approximate these curves by linear segments. Refer to the curves below for nomenclature[4,7].

Emitter Input Characteristic



Current Transfer Characteristic

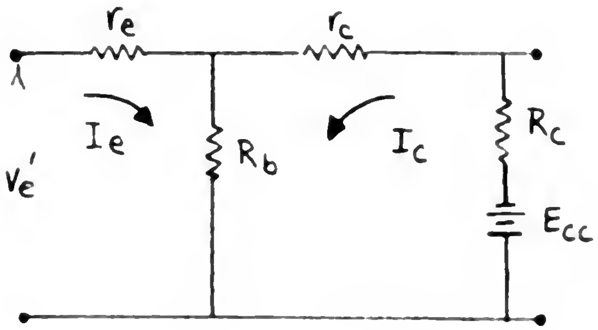


The straight line approximations to these curves can be obtained in terms of the transistor and circuit parameters

by the use of appropriate linear equivalent circuits for each region[1,4,7]. In the following development $r_b + R_b$ is replaced by R_b since in practical switching circuits $R_b \gg r_b$.

3. Cut-off Region.

Here I_e is negative; hence, it has virtually no effect on I_c. I_c remains nearly constant at its minimum value. The equivalent circuit is therefore:



$$V_e' \approx I_e r_e - \frac{E_{cc}}{R_c + r_c + R_b} \times R_b \tag{40}$$

When $I_e = 0$,

$$V_e' = -\frac{E_{cc} R_b}{R_c + r_c + R_b} \tag{41}$$

The slope of this line is

$$\frac{\partial V_e'}{\partial I_e} \approx r_e \tag{42}$$

Thus the cut-off segment is determined.

For the corresponding segment of the current transfer characteristic

$$I_c = \frac{-E_{cc} - I_e R_b}{R_b + r_c + R_c} \tag{43}$$

When $I_e = 0$,

$$I_c = -\frac{E_{cc}}{R_b + r_c + R_c} \tag{44}$$

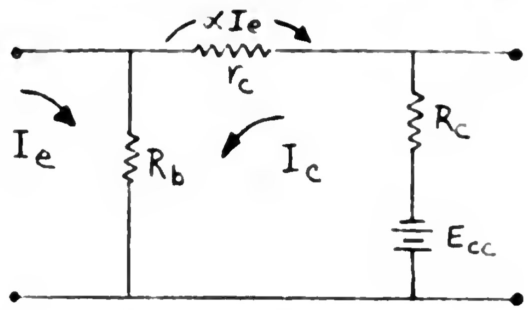
and the slope is

$$\frac{\partial I_c}{\partial I_e} = -\frac{R_b}{R_b + r_c + R_c} \tag{45}$$

The negative sign of the slope is the natural result of the opposite polarities of emitter and collector biases.

4. Transient Region.

In this region I_e is positive, hence $r_c \ll R_b$ and R_c and there is current gain ($\alpha > 1$). Using the current generator equivalent circuit we have



Writing the loop equations

$$R_b I_e \quad + \quad R_b I_c = V_e' \tag{46}$$

$$(R_b + r_m) I_e + (R_b + r_c + R_c) I_c = -E_{cc} \tag{47}$$

Then

$$I_c = -\frac{E_{cc}}{R_b + r_c + R_c} - \frac{(r_m + R_b) I_e}{R_b + r_c + R_c} \tag{48}$$

hence,

$$V_e' = \left[R_b - \frac{R_b(r_m - R_b)}{R_b + r_c + R_c}\right] I_e - \frac{R_b}{R_b + r_c + R_c} E_{cc} \tag{49}$$

$$= R_b\left[\frac{R_b + r_c + R_c - \alpha r_c - R_b}{R_b + r_c + R_c}\right] I_e - \frac{R_b E_{cc}}{R_b + r_c + R_c} \tag{50}$$

and finally,

$$V_e' = \frac{R_b\left[R_c + r_c(1-\alpha)\right]}{R_b + r_c + R_c} I_e - \frac{R_b E_{cc}}{R_b + r_c + R_c} \tag{51}$$

When $I_e = 0$,

$$V_e' = -\frac{R_b E_{cc}}{R_b + r_c + R_c} \tag{52}$$

The slope is,

$$\frac{\partial V_e'}{\partial I_e} = \frac{R_b\left[R_c + r_c(1-\alpha)\right]}{R_b + r_c + R_c} \tag{53}$$

For the current transfer characteristic the loop equations give

$$I_c = \frac{-E_{cc}}{R_b + r_c + R_c} - \frac{R_b + r_m}{R_b + r_c + R_c} I_e \tag{54}$$

$$= -\frac{E_{cc}}{R_b + r_c + R_c} - \frac{R_b + \alpha r_c}{R_b + r_c + R_c} I_e \tag{55}$$

and when $I_e = 0$,

$$I_c = -\frac{E_{cc}}{R_b + r_c + R_c} \tag{56}$$

The slope of this segment is

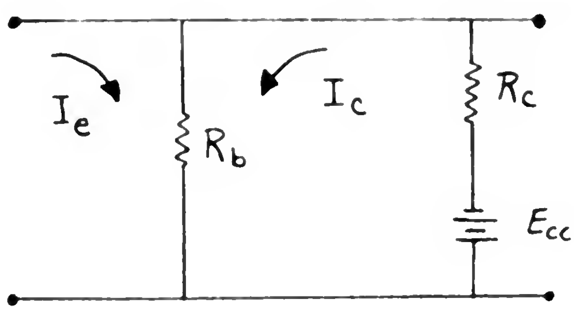
$$\frac{\partial I_c}{\partial I_e} = -\frac{R_b + \alpha r_c}{R_b + r_c + R_c} \tag{57}$$

The slopes and the intersection points at zero emitter current have now been determined for the cut off and the transient regions.

5. Saturation Region.

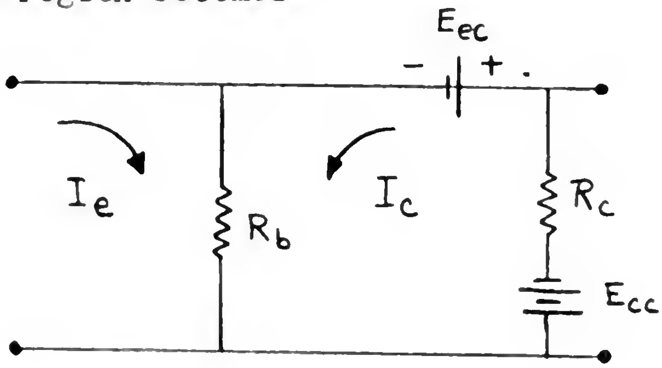
For the saturation region the emitter input segment is determined by its slope and its zero emitter voltage intercept. The corresponding current transfer characteristic is determined from its slope and its projected intercept at zero emitter current.

In this region the current gain is zero(7,9) and the back resistance of the collector, r_c , is small compared to R_c . Therefore the equivalent circuit is



Experiment has demonstrated(7) the existence of a small positive constant potential, E_{ec} between emitter and collector throughout the saturation region. Including this effect for a better approximation the equivalent circuit for the

saturation region becomes



$$V_e' = \frac{R_b R_c}{R_b + R_c} I_e - \frac{R_b}{R_b + R_c}\left(E_{cc} - E_{ec}\right) \tag{58}$$

When $V_e' = 0$,

$$I_e = \frac{E_{cc} - E_{ec}}{R_c} \tag{59}$$

The slope of this line is

$$\frac{\partial V_e'}{\partial I_e} = \frac{R_b R_c}{R_b + R_c} \tag{60}$$

For the current transfer line

$$I_c = -\frac{R_b}{R_b + R_c} I_e - \frac{E_{cc} - E_{ec}}{R_b + R_c} \tag{61}$$

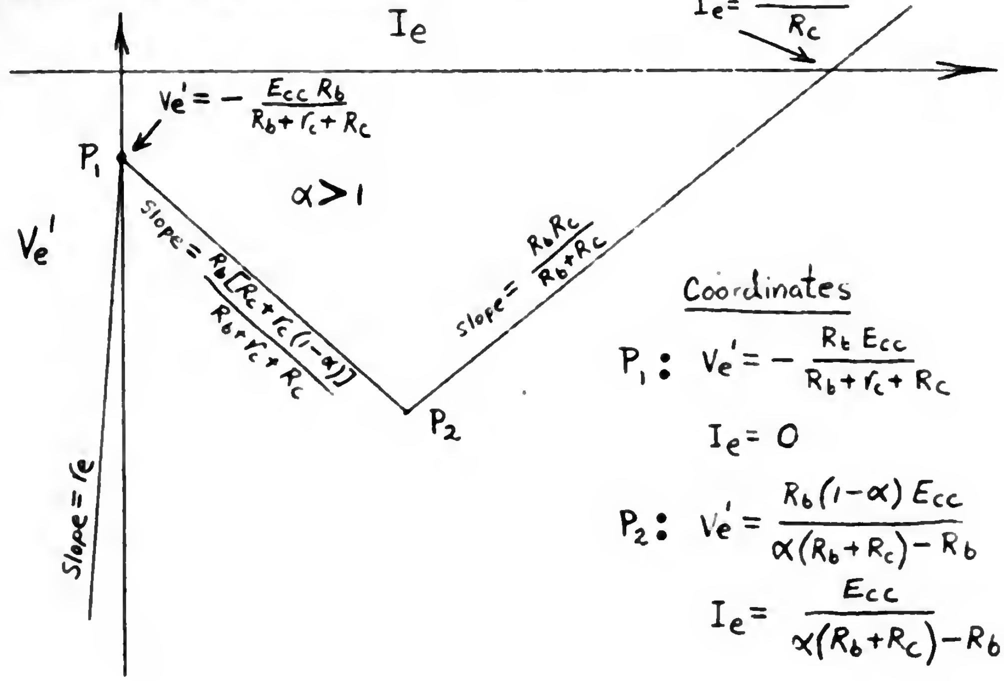
and when $I_e = 0$,

$$I_c = -\frac{E_{cc} - E_{ec}}{R_b + R_c} \tag{62}$$

and the slope is

$$\frac{\partial I_c}{\partial I_e} = -\frac{R_b}{R_b + R_c} \tag{63}$$

6. Graphical Summary.



Coordinates

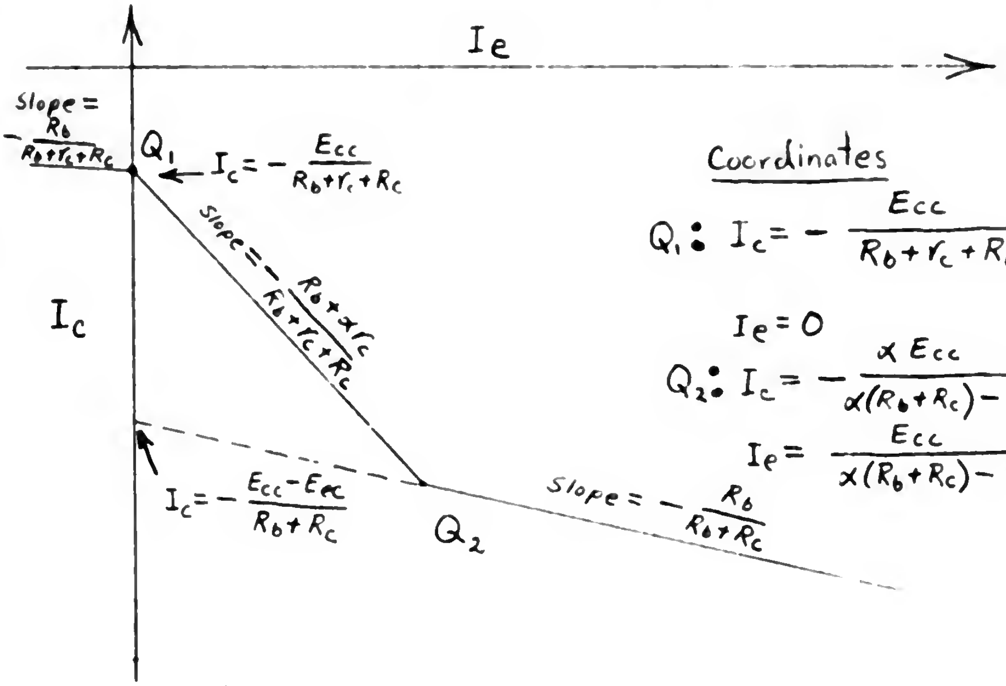
$$P_1: \quad V_e' = -\frac{R_b E_{cc}}{R_b + r_c + R_c} \tag{64}$$

$$I_e = 0 \tag{65}$$

$$P_2: \quad V_e' = \frac{R_b(1-\alpha) E_{cc}}{\alpha(R_b + R_c) - R_b} \tag{66}$$

$$I_e = \frac{E_{cc}}{\alpha(R_b + R_c) - R_b} \tag{67}$$

Emitter Input Characteristic



Coordinates

$$Q_1: \quad I_c = -\frac{E_{cc}}{R_b + r_c + R_c} \tag{68}$$

$$I_e = 0 \tag{69}$$

$$Q_2: \quad I_c = -\frac{\alpha E_{cc}}{\alpha(R_b + R_c) - R_b} \tag{70}$$

$$I_e = \frac{E_{cc}}{\alpha(R_b + R_c) - R_b} \tag{71}$$

Current Transfer Characteristic

7. Load Effects[4,7,9,22].

The foregoing analysis is based on the simple grounded base circuit in which the emitter displays the negative resistance characteristic. The mode of operation is therefore determined by the emitter load. If the load is a pure resistance intersecting the curve once in each region then operation is bistable[4,7,9]. If it is a resistance greater in magnitude than the negative resistance and intersecting only the negative resistance part of the curve, and shunted by a capacitor or an open circuited transmission line, the circuit is monostable[4,7,9]. The stable point is of course at zero emitter current.

The wave forms of the output are obtained from the current transfer curves after the form of the emitter current has been determined.[7]

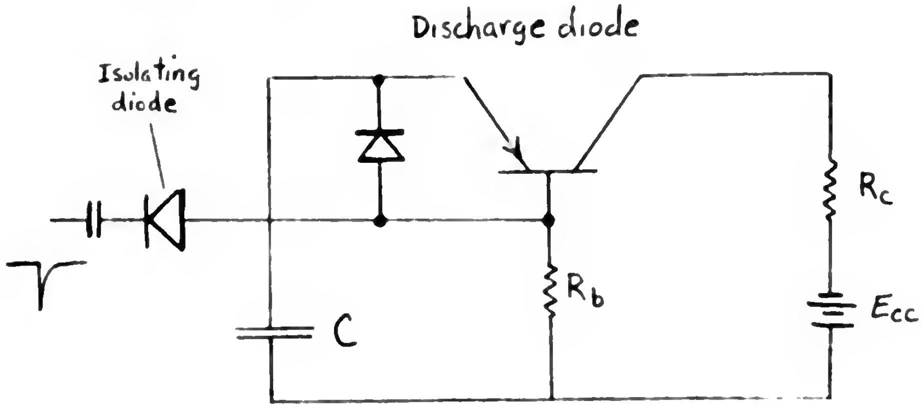
Various bias arrangements have been devised to control trigger sensitivity and to take advantage of direct coupling possibilities. For example, the V_e' $-I_e$ curve may be shifted vertically by obtaining collector bias from an E_{bb} instead of an E_{cc}[7]. A negative current bias supplied through a very high resistance will move the V_e' $-I_e$ curve horizontally, decreasing trigger sensitivity for monostable and bistable operation[7].

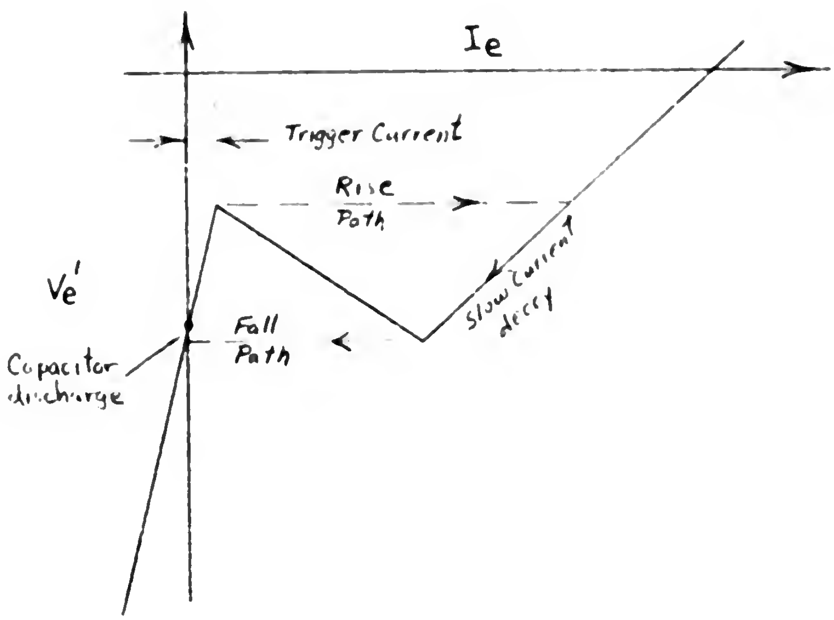
The use of semiconductor diodes as loads, isolating elements, and capacitor discharge elements gives these

basic circuits wide switching applications.

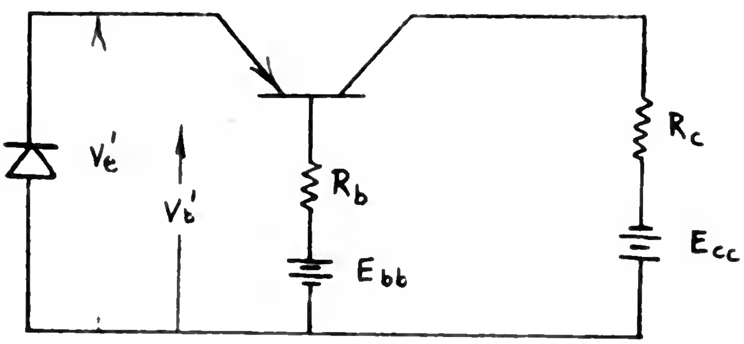
In the interests of completeness some practical examples of transistor negative resistance circuits follow.

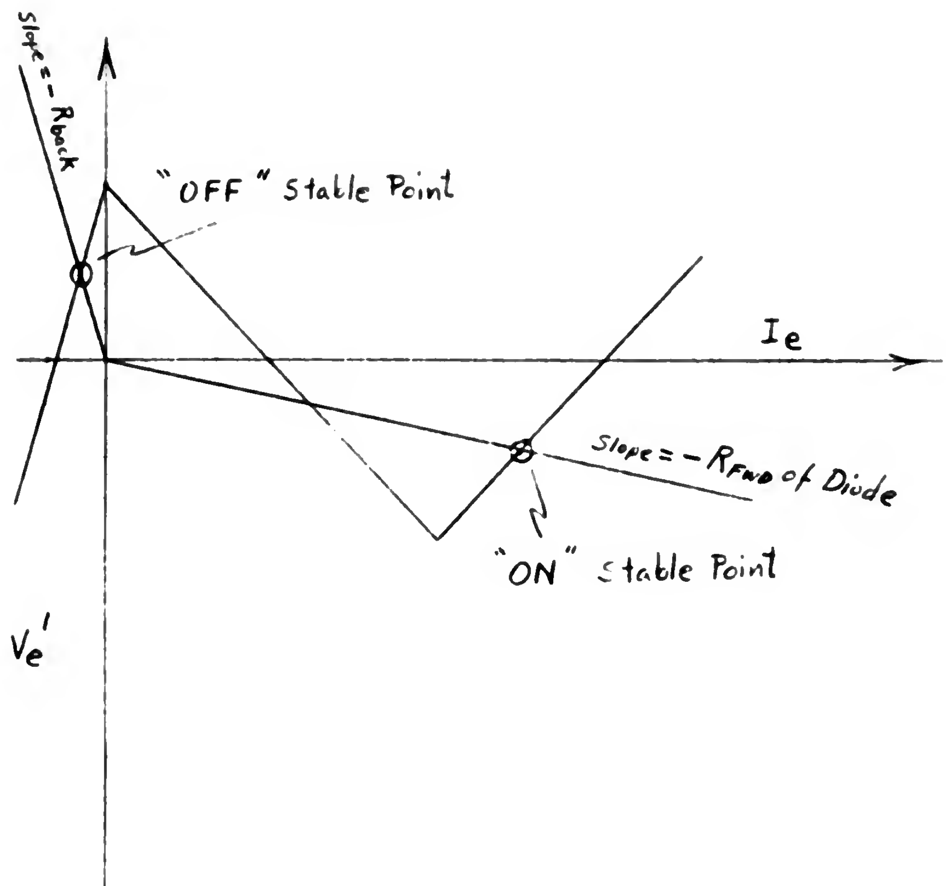
8. Practical Monostable Circuit[7].



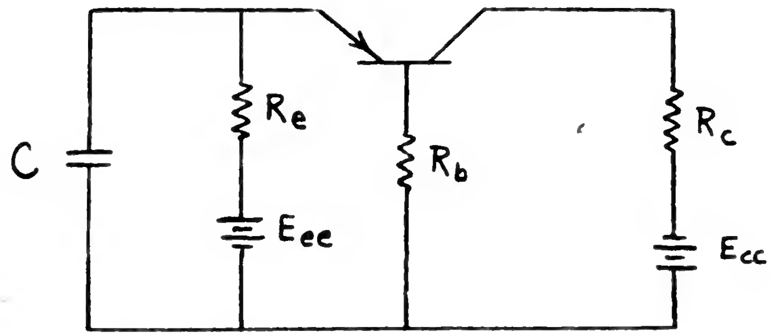


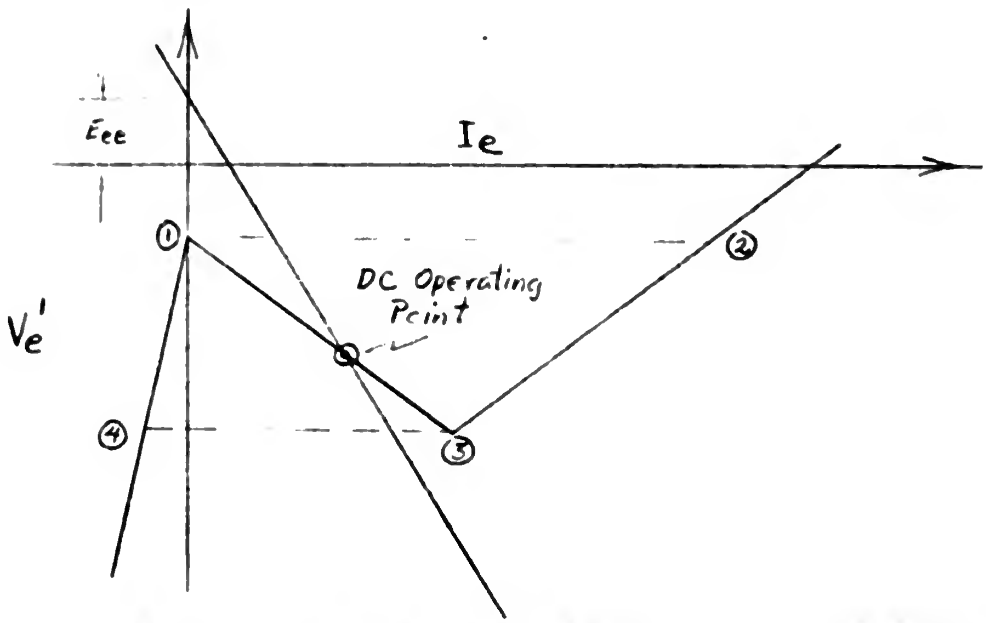
9. Practical Bistable Circuit[7].





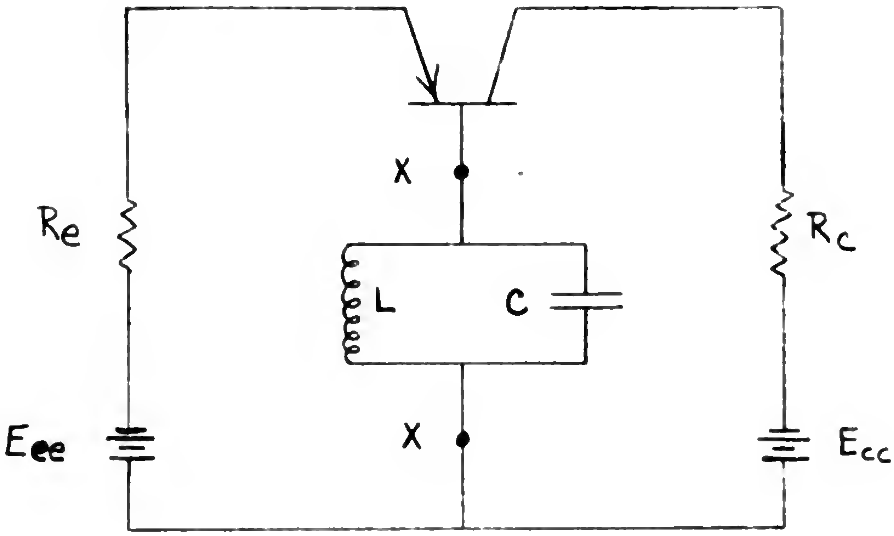
10. Practical Astable Circuit[1,7].





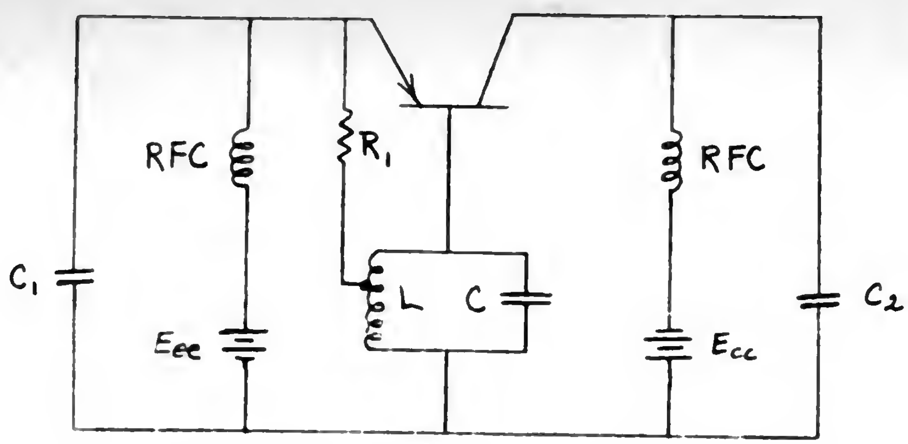
In the circuit above the operating point is set by E_{ee} or R_e or both. The operating path is 1, 2, 3, 4. The general requirements for astable operation are: a dc operating point on the transient segment, and a reactive load. A capacitor load gives multivibrator action and a series LC load will, under correct adjustment, give sinusoidal oscillations.

11. Sinusoidal Oscillators[7,22].

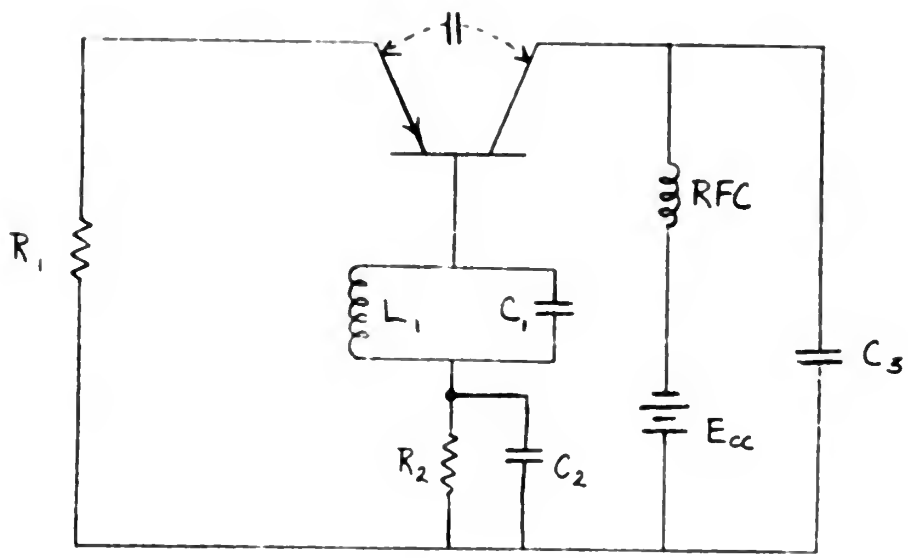


The terminals x-x display an open circuit unstable negative resistance before insertion of the parallel LC combination. The negative resistance must of course be greater than the shunt resistance of the LC load.

The next two circuits are suitable as RF oscillators. The transistors used in these circuits must be selected for high f_c of α. (See pp 31)



In the circuit above C_1 and C_2 are phase compensating capacitors. They serve to extend the high frequency limit. R_1 is a positive feedback element which may be omitted with some transistors.



R_2C_2 is a self bias network.

CHAPTER VII

FREQUENCY RESPONSE[2,11,21,23,27,29]

1. General.

The foregoing derivations have been made on the assumption that all the transistor parameters are pure resistances. This is a fair approximation over the range of usual application[14,22].

Actually r_e, r_b, and r_c, are practically independent of frequency[2], but there is a stray capacitance shunting r_c[26,27,29]. This becomes important at higher frequencies, particularly when working into a high impedance load[14]. More important is the frequency dependence of r [2,14]. r_m rolls off to about -9 db/oct above a frequency called the cut-off frequency of α [21]. It is the frequency at which the current gain is down 3 db from its low frequency value. f_c is a function of point spacing (base thickness in the case of JT)[22,27,29,30], type of minority carrier, temperature, collector voltage, and some yet undertermined variables.

For general prediction of frequency response of various resistance terminated circuits it is convenient to assume that α has a characteristic identical to an RC integrating or phase lag circuit[11,14,23,29]. Such a circuit has a response which is down 3 db at the cut off frequency and has a high frequency asymptotic slope of -6 db/oct. The frequency dependence of α is thus

$$\alpha = \frac{\alpha_0}{1 + j\frac{f}{f_c}} \tag{72}$$

where α_0 is the low frequency value of current gain.

The α of some transistors actually approachs this characteristic, but more often the high frequency asymptote is several db/oct steeper and the associated phase lag is somewhat greater(14).

From equation (15)

$$i_L = -\frac{R_{21} e_g}{\Delta} = -\frac{e_g (r_b + r_m)}{(R_g + r_e + r_b)(R_L + r_c + r_b) - r_b(r_b + r_m)} \tag{73}$$

Dividing numerator and denominator by r_c

$$i_L = -\frac{e_g\left(\frac{r_b}{r_c} + \alpha\right)}{R_{11}\left[\left(\frac{R_L}{r_c} + \frac{r_b}{r_c} + 1\right) - \frac{r_b\left(\frac{r_b}{r_c} + \alpha\right)}{R_{11}}\right]} \tag{74}$$

Then for $R_L \ll r_c$, and $r_b \ll r_m$,

$$i_L \approx -\frac{e_g \alpha}{R_{11}\left(1 - \frac{\alpha r_b}{R_{11}}\right)} \tag{75}$$

Substituting the complex α it is seen that

$$i_L = -\frac{e_g \alpha_0}{R_{11}\left(1 + j\frac{f}{f_c}\right)\left[1 - \frac{\alpha_0 r_b}{R_{11}\left(1 + j\frac{f}{f_c}\right)}\right]} \tag{76}$$

$$= -\frac{e_g \alpha_0}{R_{11}\left[1 - \frac{\alpha_0 r_b}{R_{11}}\right] + j R_{11}\frac{f}{f_c}} \tag{77}$$

$$i_L = - \frac{e_g \alpha_o}{R_{11}\left(1 - \frac{\alpha_o r_b}{R_{11}}\right)\left(1 + j\frac{f}{\left(1 - \frac{\alpha_o r_b}{R_{11}}\right) f_c}\right)} \qquad (78)$$

Thus the load current is down 3 db when

$$f = \left(1 - \frac{\alpha_o r_b}{R_{11}}\right) f_c \qquad (79)$$

Then, under the assumptions given, the circuit cut-off frequency is $1 - \frac{\alpha_o r_b}{R_{11}}$ times the α cut-off frequency.

For the grounded base circuit used in the previous examples

$$r_b = 290$$
$$R_{11} = 1030$$
$$r_m = 33710$$
$$r_c = 18710$$

hence,

$$\alpha_o = \frac{r_m + r_b}{r_c + r_b} = \frac{R_{21}^*}{R_{22}^*} = \frac{34}{19} = 1.79 \qquad (80)$$

so that

$$1 - \frac{\alpha_o r_b}{R_{11}} = 1 - \frac{1.79 \times 290}{1030} = 1 - .5 = .5 \qquad (81)$$

This indicates that the circuit cut-off frequency will be approximately one half of the α cut-off frequency. However, the approximation $R_L \ll r_c$ is not satisfied in this

example. Therefore the previous result should be modified as follows:

$$\text{Cut-off frequency reduction factor} = 1 - \frac{\alpha_o r_b}{2 R_{11}}$$

$$= 1 - \frac{1.79 \times 290}{2060} = .75$$

Qualitatively the effect illustrated here is that positive feedback increases gain at the expense of bandwidth. This can be compensated by the appropriate application of negative feedback.

2. Physical Aspects[13,21].

Point spacing, s, of PCT or base width of JT have an effect on f_c which is similar to transit time effects in vacuum tubes[31]. The transit time in transistors also depends on the mobility of the minority carrier, μ, the resistivity of the base material, ρ, and the emitter current. For a PCT the transit time is given by

$$T = \frac{s^3 \sigma}{\mu I_e}$$

The mobility of electrons as minority carriers is about 50% greater than that of holes[19]. For this reason f_c is appreciably greater for P-base PCT and npn JT than it is for N-base PCT or pnp JT[14,10].

In addition to basic research in solid state physics the efforts to increase the frequency range of PCT are

directed toward decreasing s. There are, however, definite limitations to the practical reduction of point spacing. It has been found that for moderate values of ρ, r_b increases rapidly with decreasing s. For low values of , r increases linearly with decreasing s. It is therefore necessary that ρ be low in order to preserve stability.

The other limitations to reduced point spacing are the mechanical difficulty, and the stray capacity shunting the points[13,21].

At present the above techniques have resulted in f_c of more than 30 mc and oscillation in the 100 to 300 mc range. To achieve this, point spacings as small as one half mil were used[13,21].

CHAPTER VIII

NOISE AND TEMPERATURE EFFECTS

1. Noise in Transistor Circuits.

A considerable amount of analytical work(6,8) has been done on transistor noise, but here we will consider only the more general aspects of noise. We are particularly interested in the methods of insuring relatively low noise.

Most of the noise of transistors is generated at the collector(8). The emitter noise power is of the order of 40 db below collector noise power(8).

In current literature the noise is usually expressed as a noise figure at 1000 cps measured over an incremental bandwidth(2,6,8,9,27).

$$\text{Noise Figure (db)} = 10 \log \frac{P_L}{G_F P_G}$$

where P_L = Noise power in load for the specified bandwidth

G_F = Forward power gain

P_G = Thermal noise power in R_g for the same bandwidth

Noise figure depends on frequency, bias currents and R_g but is independent of R_L [6].

For the grounded base connection noise figure is minimum when $R_g = R_{11}{}^* = r_e + r_b$ [6,27]. This value of R_g does not, in general, match the input impedance of the circuit. A compromise must therefore be made between noise and gain. The appropriate compromise can be made with the aid of the table below (due to Montgomery [8]).

$\dfrac{R_g}{R_g\,(\text{Optimum Noise})}$	½ or 2	⅕ or 5	1/10 or 10
Increase in noise figure (db)	0.5	2.6	4.8

In the grounded emitter connection R_g for optimum noise is approximately the same as for the grounded base connection, but it is usually much smaller than that for maximum gain [8,14].

There is little difference in the noise figures obtainable with any of the fundamental amplifier circuits except that the noise figure for reverse transmission through a grounded collector circuit is likely to be much higher than the others [14].

The variation of noise power with frequency for both PCT and JT is fairly well expressed as[2,8,9,14,22,27]:

$$\text{Noise Power} \propto \frac{1}{f^n} \text{ , where } n \text{ is from } 1.1 \text{ to } 1.2$$

Noise figure increases with increasing $|V_c|$ over the normal operating range of voltage[6].

2. Temperature Effects.

The two parameters most seriously affected by temperature variation are r_c and α[9].

The collector back resistance, r_c decreases with temperature[2,3]. This implies that I_{c_0} increases with temperature. Hence, there may be serious faults in switching circuits[9], since, in switching applications, it is usually desirable that I_{c_0} be as low as possible.

The current gain, α , increases with temperature and partially offsets the decreasing r_c thus keeping the loss of small-signal gain low over the operating range of temperatures (-40°C to + 70°C)[3,9].

If the operating temperature exceeds 80°C in most of the present types there will be permanent changes in the small signal and dc parameters[3].

In general, from the stability standpoint, the effect of increasing temperature is to impair stability.

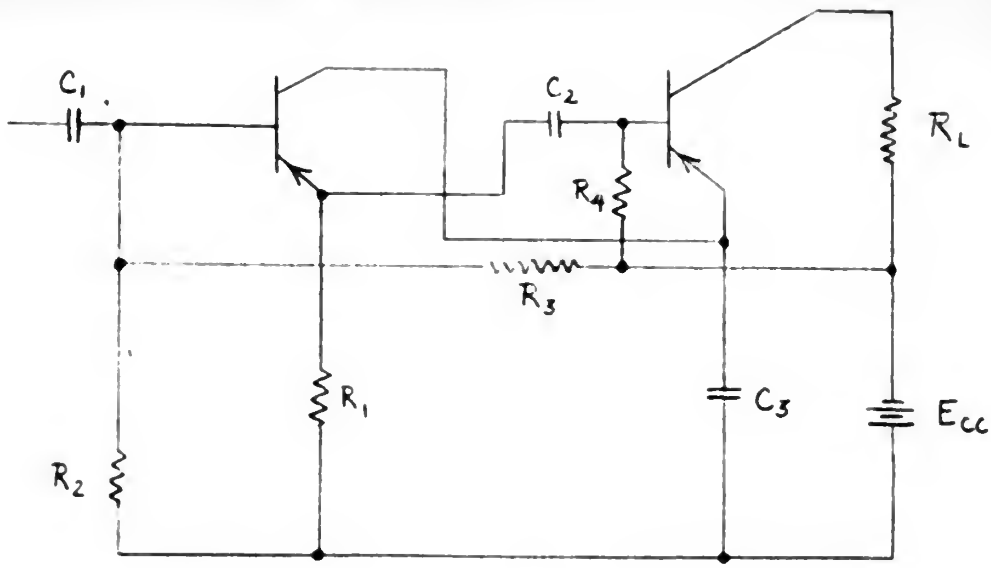
CHAPTER IX

MISCELLANEOUS CASCADED AMPLIFIER CIRCUITS

In cascading transistor amplifier stages it is important to give special consideration to interstage coupling. If the usual heavy, bulky audio transformers are used then most of the weight and space advantages of transistors are immediately sacrificed. In some applications transformer coupling will be unavoidable. If such is the case, it is possible to minimize the weight and space increase by devising circuits in which the dc component through the transformer windings is minimum[22] (preferably zero). The circuitry used for this purpose is a compromise in efficiency.

Transformers may frequently be avoided by taking advantage of the variety of impedance transformations available in the three transistor connections[22].

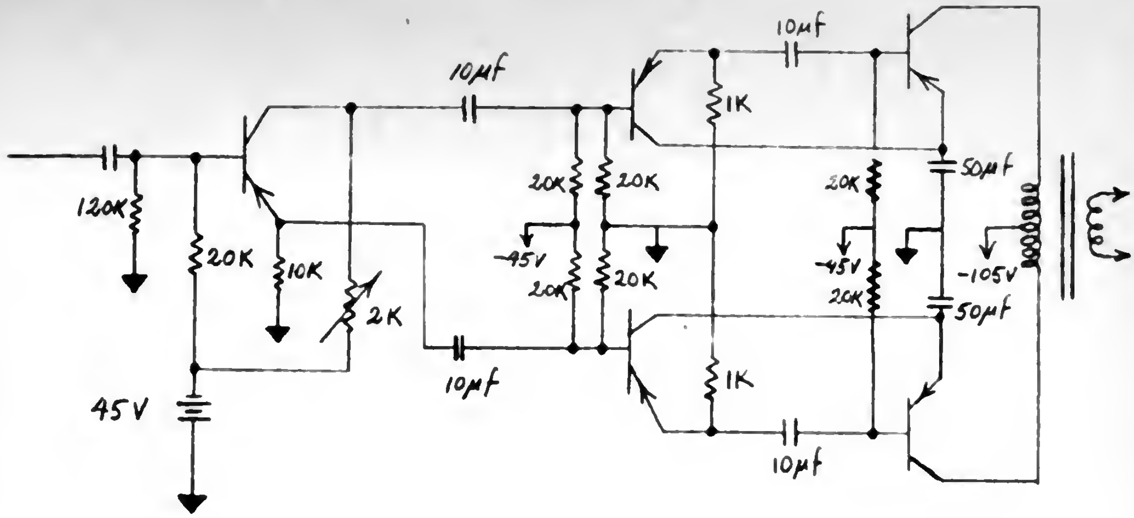
The problem of power supplies for cascaded amplifiers is another subject for compromise[16,22]. The problem is to keep the number of separate power sources to a minimum without too great a loss in overall efficiency. Some methods of dealing with the problem are illustrated in the miscellaneous cascaded amplifier circuits which follow[22].



AUDIO AMPLIFIER (Class A, single ended)[22]

In this circuit a sacrifice is made in overall efficiency to permit operation from one battery[16].

The first transistor is used at low power gain in a grounded collector connection, and supplies emitter bias for the second stage; operated grounded emitter.



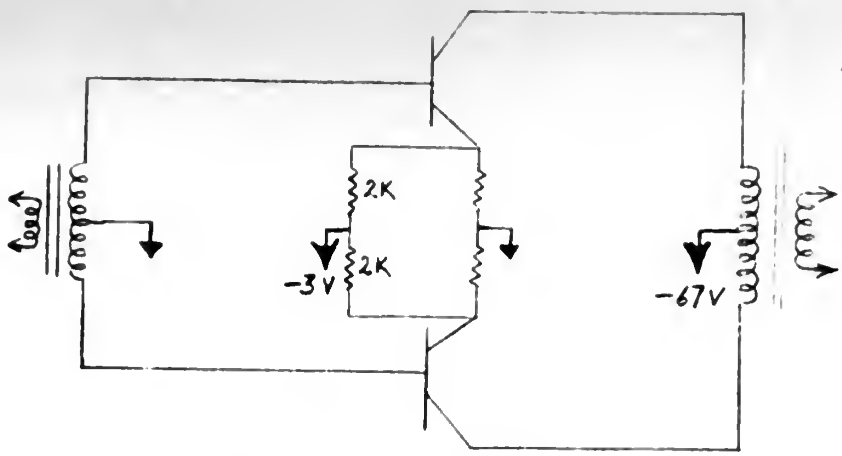
AUDIO AMPLIFIER (Class A, push-pull)

The grounded collector to grounded emitter cascade, as used above, is a common impedance matching technique(22). In a similar way, a grounded base to grounded collector connection is particularly useful for low impedance loads.

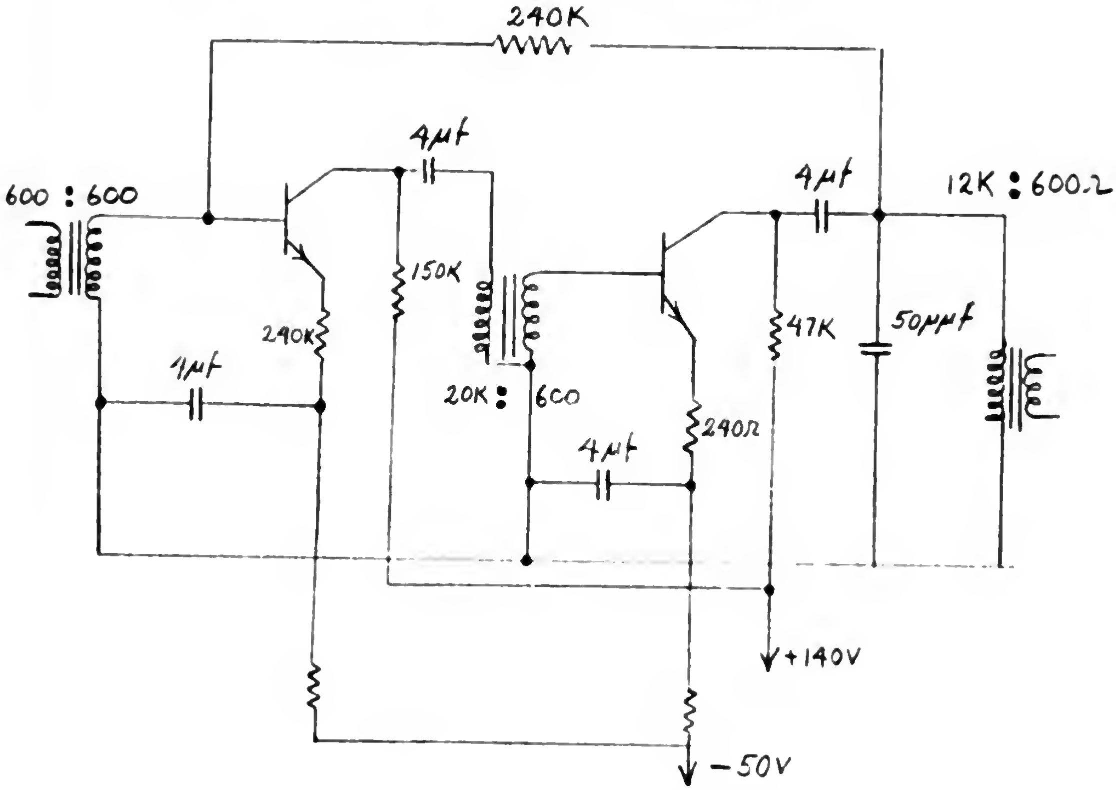
The first stage is a split load phase inverter. Its output is balanced by the 2 K variable resistance.

Note the use of the rather large capacitors. This is necessary, of course, to maintain a suitable coupling time constant and adequate bypass of low impedance circuits. Fortunately these are usually low voltage capacitors.

Another point which should be mentioned here is that where a large power output is required the emitter is not a suitable output terminal because of its low current handling ability(14).



AUDIO POWER AMPLIFIER (Class B)[22]



NPN JUNCTION TRANSISTOR AUDIO AMPLIFIER[22]

The significant feature of the circuit above is the use of resistance shunt feed to reduce the dc magnetization of the transformer core. The circuit is classified as grounded emitter but emitter degeneration is employed to improve stability.

CHAPTER X

CONCLUSIONS

In adapting transistors to perform the functions commonly performed by vacuum tubes there are two general approaches. One of these consists of duplicating, as nearly as possible, the circuitry of an existing vacuum tube circuit, adapting as necessary to satisfy the special characteristics of the transistor. For example, a transistor multivibrator can be made to operate using essentially the same circuitry as that used for the vacuum tube counterpart.

The second method is to redesign the circuit to take full advantage of the transistor characteristics. In contrast to the early work on circuit applications the recent trend is toward the second method. The circuits in this paper fall into this latter category. In the example of the multivibrator, it has been demonstrated that the transistor counterpart is more efficiently constructed with a single transistor.

At this time there exist several handicaps which seriously limit transistor performance in certain applications. The most important of these is the breakdown of performance at elevated temperatures. Another is the frequency limitation. A third difficulty is peculiar to the short circuit unstable PCT. This problem involves keeping

the external circuit impedances high, sometimes well outside the frequency range of interest. The corresponding problem in vacuum tube circuits is to keep this impedance low above the frequency range of interest. Keeping stray shunt impedances high at high frequencies may be a very difficult problem.

Because of the limitations mentioned above, the practical applications of transistors on a commercial or military basis have thus far been limited to such services as audio amplifiers, audio oscillators, and pulse shaping and trigger circuits. Further extension of the fields of reliable application is contingent to some extent on improved production techniques, but to a greater extent, on the progress of basic research in the supporting solid state physics.

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